DESIGNING WITH THE μA703 MONOLITHIC RF-IF AMPLIFIER

by G. J. Estep

INTRODUCTION

The IF-RF amplifiers and mixers in most receiving systems generally employ few components in addition to the tuned circuits. Since tuned circuits cannot be economically integrated using existing technology, use of the remaining components in integrated form will be worthwhile only if a substantial improvement in performance or stability can be obtained, or if simplified design procedures can be used.

These improvements are provided by the emitter-coupled amplifier, the μA703, shown in Fig. 1. Designed for use with transformers and/or tuned circuits, the μA703 eliminates the biasing resistors, bypass capacitors, and coupling capacitors required in conventional amplifiers. Operation is essentially unilateral, and the circuit has a symmetrical output current-limiting characteristic. This provides increased available power gain, simplified tuning, and superior limiting performance which greatly reduces the de-tuning, phase shift, and blocking that occur when conventional amplifiers saturate. The device may also be used as a harmonic mixer and is especially attractive when used as an FM limiter.

The biasing arrangement takes advantage of the closely-matched characteristics of integrated circuit transistors, while the dc return path for biasing the input and output is provided by interstage transformer coupling. A single voltage supply is all that is required for operation.

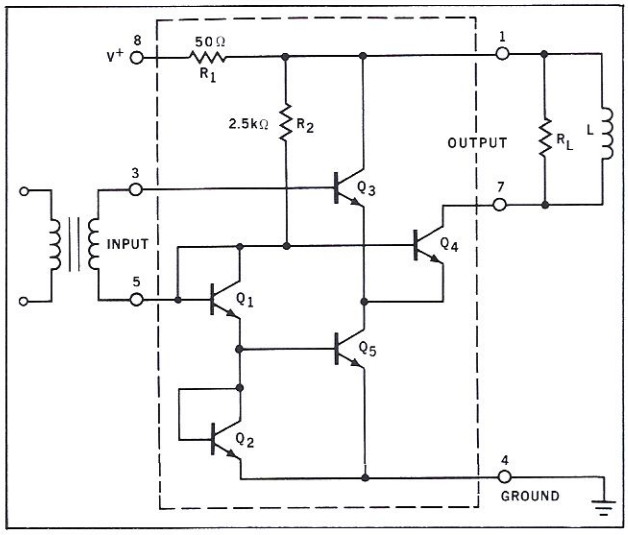

Fig. 1. Schematic diagram of the μA703.

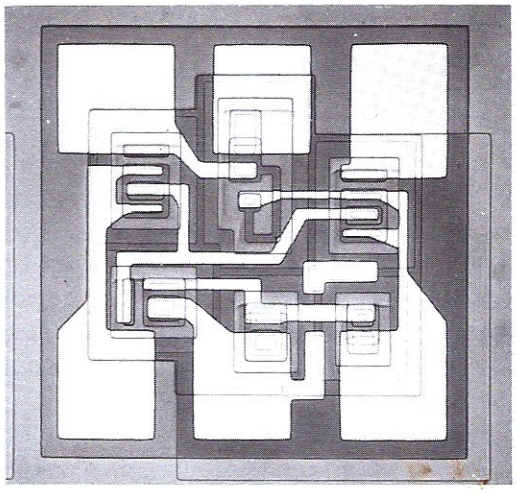

Fig. 2. Photomicrograph of the μA703.

The simplicity of the amplifier is demonstrated by the fact that it is constructed on a 20-mil-square silicon die (Fig. 2) -- a size which is small by comparison to many discrete transistors.

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Circuit Description

The dc biasing of the amplifier is easily understood, if it is assumed that all parts within the circuit are well matched and the transistor current gains are high enough that the base currents can be neglected. It is also assumed that the transformer windings, particularly in the input circuit, have negligible dc resistance.

The collector current of the biasing transistor, Q_2, is

$$I_{C2} = \frac{V^+ - 2V_{BE}}{R_2 + 2R_1} \quad . \tag{1}$$

Since $V^+ \gg 2V_{BE}$ and $R_2 \gg 2R_1$, this current is substantially independent of the transistor characteristics and is simply a function of the supply voltage and a single resistor.

The transistors are assumed to be identical, hence the collector current of the current source, Q_5, is equal to that of Q_2 because their bases are fed from a common voltage point.[1] The collector current of Q_5 splits evenly between Q_3 and Q_4 with zero input signal. When the amplifier is driven into limiting, this current is alternately switched between Q_3 and Q_4. To prevent saturation of Q_4 when the amplifier is driven with a large signal, the load resistance must be low enough that current limiting occurs before the output voltage drops to $2V_{BE}$. Thus, for a transformer-coupled output,

$$R_L \leq \frac{2(V^+ - 2V_{BE})}{I_{C2}} \leq 2R_2 \tag{2}$$

A more detailed analysis of the μA703 transfer function is given in Appendix I, along with some qualitative discussion of temperature and frequency effects. Curves of the two-port parameters are included to illustrate typical device characteristics.

RF Amplifier

Unneutralized small-signal amplifiers have a feedback path which can cause the real part of the input admittance of the device to become negative for certain combinations of frequency and load impedance. If this situation occurs and the source admittance is equal to or less than the negative real part of the input admittance, self-sustaining oscillations are possible. For the typical amplifier, the designer has one of two choices: feedback that is equal in magnitude and opposite in phase to that inherent in his device can be applied around the circuit, or the interstage coupling networks can be loaded sufficiently to make the input admittance always greater than zero. These solutions have major drawbacks -- neutralization is at best a marginal solution since it is very frequency-sensitive and must be different for each situation. Loading the interstages is less frequency-sensitive, but the resultant mismatch loss is not always tolerable. The most satisfactory solution would be to use a device which possesses negligible intrinsic feedback -- if something of that description does indeed exist. For the μA703, the internal feedback is an order of magnitude less than most high-frequency transistors. For example, a reverse transadmittance of less than 0.001 mmho at 10MHz is not unusual.

The effect of input and output loading on the stability of the amplifier can be best seen from the expression for power gain of the device. Characterizing the device as an active

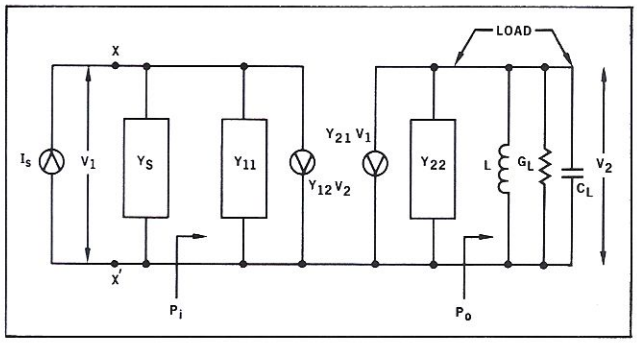

Fig. 3. Tuned amplifier equivalent circuit.

two-port and using short-circuit admittance parameters to describe its behavior gives the equivalent circuit shown in Fig. 3, where the following y-parameters have been used:

y_{11} - input admittance, $g_{11} + jb_{11}$

y_{12} - reverse transadmittance, $g_{12} + jb_{12}$

y_{22} - output admittance, $g_{22} + jb_{22}$

y_{21} - forward transadmittance, $g_{21} + jb_{21}$

Y_L - load admittance, $g_L + jb_L$

Y_S - source admittance, $g_s + jb_s$.

The expression for the input power for the circuit of Fig. 3 is:

$$P_i = |V_1|^2 \, R_e \left[Y_{in}\right] = |V_1|^2 \, R_e \left[y_{11} - \frac{y_{12}y_{21}}{y_{22} + Y_L}\right] \tag{3}$$

and the output power is:

$$P_o = |V_2|^2 \, R_e \left[Y_L\right] \quad . \tag{4}$$

From Fig. 3,

$$|V_2| = \frac{|y_{21}| \, |V_1|}{|y_{22} + Y_L|} \tag{5}$$

So

$$P_o = \frac{|y_{21}|^2 \, R_e \left[Y_L\right]}{|y_{22} + Y_L|^2} \, |V_1|^2 , \tag{6}$$

and the power gain becomes

$$\frac{P_o}{P_i} = \frac{\left|y_{21}\right|^2 R_e \left[Y_L\right]}{\left|y_{22} + Y_L\right|^2 R_e \left[y_{11} - \frac{y_{12}y_{21}}{y_{22} + Y_L}\right]} \quad (7)$$

It can be seen from Equation 7 that the amplifier will be stable if

$$R_e \left[y_{11} - \frac{y_{12}y_{21}}{y_{22} + Y_L}\right] > 0. \quad (8)$$

Equation 8 can be manipulated to yield[2, 3]

$$2g_{11}(g_{22} + g_L) - R_e \left[y_{12}y_{21}\right] - \left|y_{12}y_{21}\right| > 0. \quad (9)$$

If the power gain is reckoned from the terminals XX' (Fig 3), Equation 9 becomes

$$2(g_{11} + g_s)(g_{22} + g_L) - R_e \left[y_{12}y_{21}\right] - \left|y_{12}y_{21}\right| > 0. \quad (10)$$

Inspection of the above equation reveals that, for unconditional stability at all frequencies for any combination of source and load conductance, y_{12}, the reverse transadmittance, must equal zero. Since in practice y_{12} is finite, the source and load conductances must be manipulated to satisfy the stability criterion.

If it is assumed in Equation 9 that the real part of y_{12} is nearly zero, an expression can be obtained which will lead to a useful figure of merit for a two-port.

If $R_e \left[y_{12}y_{21}\right]$ approaches zero, Equation 9 becomes

$$2g_{11}(g_{22} + g_L) - \left|y_{12}y_{21}\right| \geq 0. \quad (11)$$

which for conjugate matching becomes

$$4g_{11}g_{22} \geq \left|y_{12}y_{21}\right|. \quad (12)$$

For conditional stability,

$$4g_{11}g_{22} = \left|y_{12}y_{21}\right|. \quad (13)$$

The maximum gain which can be obtained from the two-port when the device is neutralized and conjugately matched is, from Equation 7

$$\frac{P_o}{P_i} = \frac{\left|y_{21}\right|^2}{4g_{11}g_{22}}. \quad (14)$$

Using Equation 13, the maximum stable gain will be

$$GMS = \left|\frac{y_{21}}{y_{12}}\right|. \quad (15)$$

Obviously, the smaller y_{12} is, the greater will be the potentially available stable gain. For very small values of y_{12}, the gain will be determined essentially by the input and output parameters of the device.[4] For typical high frequency transistors at 10 MHz, the GMS is 20 decibels, while a GMS of 40 decibels can be obtained for the integrated emitter-coupled amplifier. Thus, fewer stages are required to achieve a given total gain.

A simplified design procedure for a 30-MHz amplifier can now be presented which will serve to illustrate the practical aspects of the preceding discussion.

The first problem that must be considered is: "How much gain can be realized at a given frequency?" To evaluate this problem, consider Equation 15 and the device parameters listed in Table I (taken from the curves of Appendix I). At 30 MHz, the forward transadmittance, y_{21}, is equal to 32 mmho and the reverse transadmittance, y_{12}, is 0.004 mmho. The maximum stable gain, GMS, is therefore 39 decibels.

TABLE I.

DEVICE PARAMETERS.

PARAMETER	CONDITION	VALUE	UNITS
Input Resistance	V^+ = 12V, f = 30 MHz	1.7	kΩ
Input Capacitance	V^+ = 12V, f = 30 MHz	9.0	pF
Output Conductance	V^+ = 12V, f = 30 MHz	0.08	mmho
Output Capacitance	V^+ = 12V, f = 30 MHz	2.5	pF
Forward Transadmittance	V^+ = 12V, f = 30 MHz	32/140°	mmho
Reverse Transadmittance	V^+ = 12V, f = 30 MHz	0.004	mmho

To obtain this gain, two conditions must be satisfied by the circuit. First, the source and load admittances have to be conjugately matched to the device: otherwise maximum transfer of power from source to load cannot be obtained. Second, the layout of the components must be such that feedback from output to input is minimized.

The turns ratio of the input and output transformers are selected to provide a conjugate match to the device

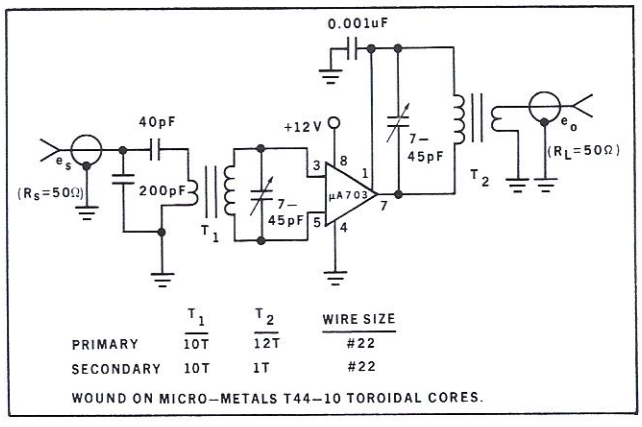

Fig. 4. The μA703 RF amplifier.

TABLE II.

AMPLIFIER PERFORMANCE.

PARAMETER	CONDITION	VALUE	UNITS
Power Gain	$V^+ = 12V$, f = 30 MHz	35	dB
Bandwidth	$V^+ = 12V$, f = 30 MHz	1	MHz
Noise Figure	$V^+ = 12V$, f = 30 MHz	6	dB
Power Gain (calculated)	Eq. 14	37.4	dB
Maximum Stable Gain	Eq. 7	39	dB

parameters listed in Table I. The actual circuit details are shown in Fig. 4, and the performance is listed in Table II. Excellent agreement is seen to exist between measured and calculated power gain. Interaction between interstage coupling networks is minimal.

Note that the amplifier does not utilize the limiting characteristic of the emitter-coupled pair and will saturate for large output swings. The input impedance does not drop, however, as the amplifier is driven into saturation.

Limiting RF Amplifier

The preceding discussion of the small-signal amplifier can be extended to limiting amplifiers without exception. The only additional requirement that must be fulfilled is that given by Equation 2, which ensures that the output transistor will not be saturated.

An amplifier which has a symmetrical output-voltage limiting characteristic can be obtained from the circuit of Fig. 4 by a minor modification in the output transformer. The effective turns ratio is adjusted so that the 50-ohm load, in parallel with the output resistance of the amplifier, causes the inequality of Equation 2 to be satisfied.

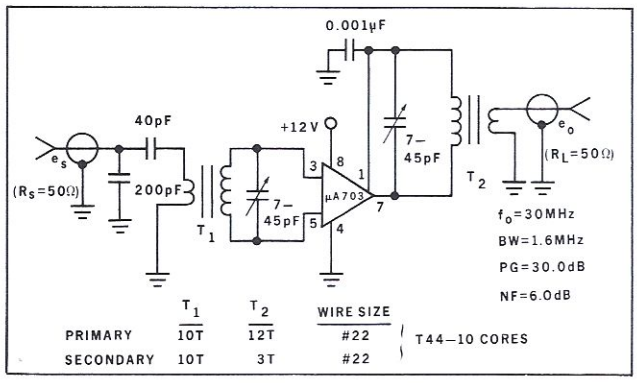

Fig. 5. Limiting RF amplifier.

For a 12-volt power supply and a 5-mA maximum collector current, the effective load should be less than 4.5 kilohms. Figure 5 shows the modified amplifier, designed for a 4500-ohm load and an output swing of 22 V_{pp}, with the measured performance characteristics.

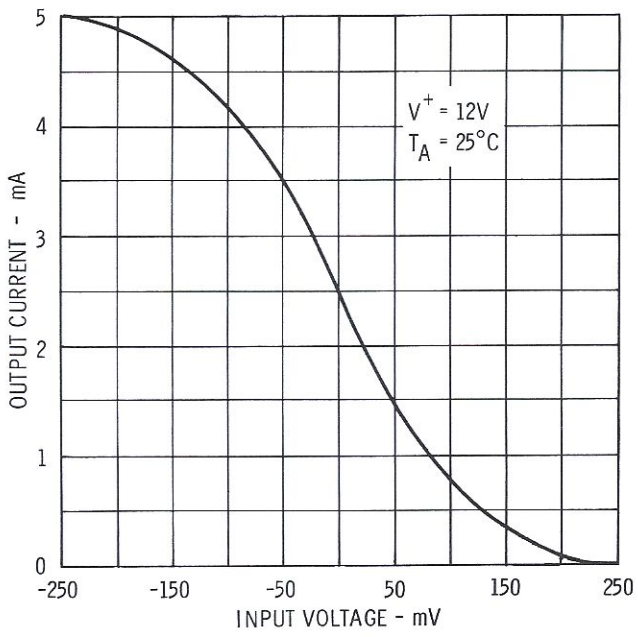

Fig. 6. Transfer function.

The current-limiting characteristic of the device is shown in Fig. 6. Fig. 7 shows the effect of large input signals on the input parameters.

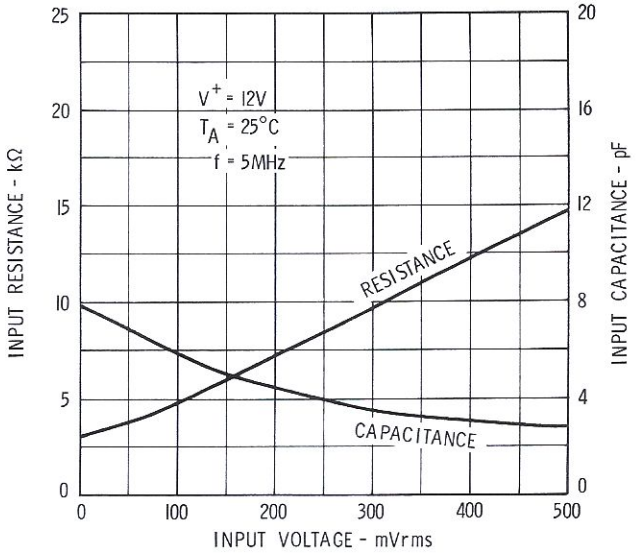

Fig. 7. Input parameters.

Oscillator

The design of transistor oscillators is complicated by the loading of the tank circuit caused by the decrease in input impedance of the device under large-signal conditions. The

usual solution is to provide some sort of impedance isolation between input and output, such as high turns ratio transformers or emitter followers in the feedback loop.

The input impedance of the μA703, however, tends to increase with large-signal inputs, as can be seen from Fig. 7. This simplifies the oscillator design, because the effect of input parameter variations can be neglected (depending upon the input voltage swing and becoming more valid for larger signals, but usually permissible for effective loads of 2 kilohms of less).

Predictable oscillator performance can be obtained by utilizing the Barkhausen criterion for oscillations, which states that the necessary and sufficient condition for oscillations to occur is:

$$1 - GH = 0, \tag{16}$$

where GH is the loop gain of the system; G is the forward transfer function, and H is the reverse (feedback) transfer function.

For the simplified equivalent circuit of Fig. 3 (assuming that the input admittance is small compared to the load admittance and the oscillator frequency is less than the corner frequency of the forward transadmittance), the forward transfer function is given by

$$G = \frac{V_2}{V_1}$$

$$= \frac{y_{21}}{g'} \quad \frac{1}{1 - j\omega L g' \left[1 - \frac{\omega^2}{\omega_o^2}\right]}, \tag{17}$$

where

$$\omega_o^2 = \frac{1}{LC'}$$

and C' = load capacitance, C_L, plus amplifier output capacitance, C_o

$$g' = g_L + g_{22}$$

Using a feedback transformer with a 1:1 turn ratio and assuming that the coefficient of coupling is unity (which is usually justified if a toroidal core is used), the reverse transfer function, H, will be equal to 1. Therefore, for oscillations to occur at $\omega = \omega_o$, the load conductance must be small enough that

$$\frac{y_{21}}{g'} \geq 1. \tag{18}$$

The effect of the phase shift contributed by the high frequency pole of y_{21} is to reduce the actual frequency of oscillation but, since in most tuned oscillators, a trimmer is used to set the final frequency, a more exact analysis is not justified. The effect of variations in input parameters was assumed negligible in the preceding discussion and any exact analysis must account for them, as well as the transformer coupling coefficient and losses.

The magnitude or the output voltage can be predicted quite well if the total voltage swing at the output collector of the device is allowed to be no more than 5 V_{pp} (since for a 1:1 transformer, the maximum input voltage rating of the device will be exceeded and the emitter-base junctions may be damaged), and the effective collector load resistor is

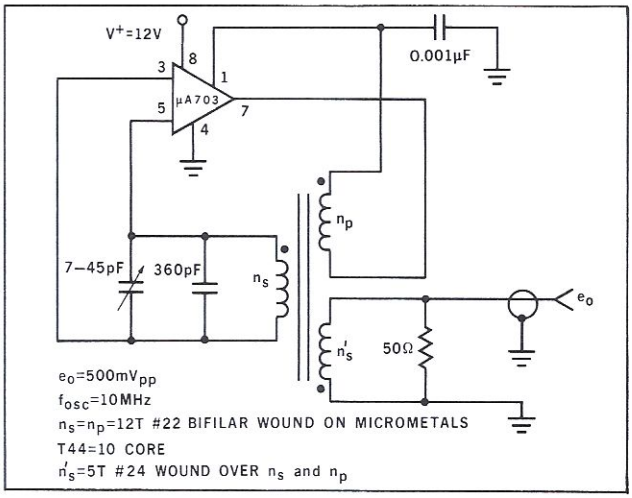

Fig. 8. 10-MHz oscillator.

small enough to ensure that current limiting occurs, but large enough that oscillations are maintained. The value of the load resistor can be determined from the expression

$$R_L = \frac{V_{pp}\ (\text{allowed})}{\text{max p-p output current}}. \tag{19}$$

An oscillator was designed using the preceding criteria with the results shown in Fig. 8. The circuit was designed for a collector voltage swing of 2 V_{pp}, hence, from Equation 19, an effective load resistance of 400 ohms is required, using the typical maximum output current of 5 mA. Figure 9 shows an oscilloscope photo of the output waveform, and the frequency stability as a function of power supply appears in Fig. 10.

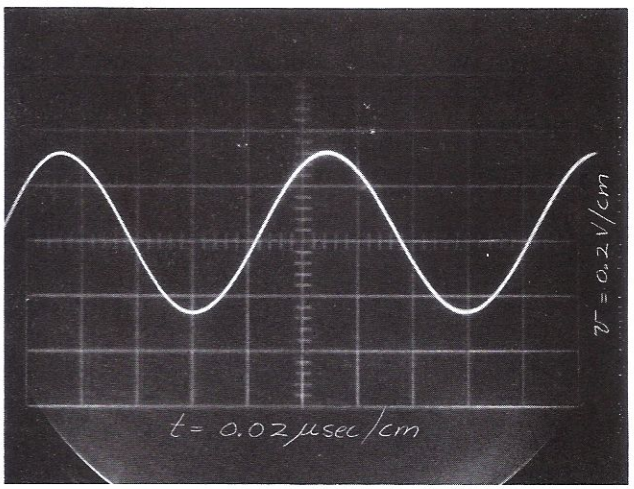

Fig. 9. Oscillator output waveform.

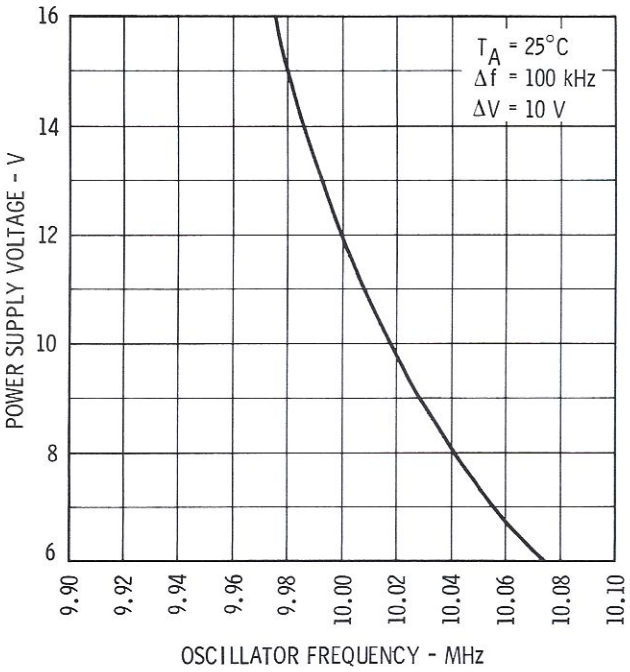

Fig. 10. Frequency stability of 10-MHz oscillator.

Harmonic Mixer

The transfer function of the μA703 is given by the function (discussed in Appendix I).

$$y_{21}(0) = -\frac{qI_{C4}}{kT}\left[1 - \frac{I_{C4}}{I_T}\right] \qquad (20)$$

This function describes the values of the small-signal transconductance for a specified source-to-output current ratio. What we would like to determine now is, which function will best describe the time varying transconductance, and how does the output compare to a conventional mixer? The answer to these questions may be found by substituting the expression for I_{C4} as a function of time into Equation 20. From Equation A-8 in Appendix I, the small-signal output current may be written,

$$i_{C4}(t) = io(t)$$

$$= I_T / (\exp qv_{in}(t)/kT + 1) \qquad (21)$$

since the source current, I_T, is constant and independent of input signal.

$$y_{21}(t) = -AI_T \frac{1}{\exp Av_{in}(t) + 1}\left(1 - \frac{1}{\exp Av_{in}(t) + 1}\right)$$

$$= -AI_T \frac{\exp Av_{in}(t)}{(\exp Av_{in}(t) + 1)^2}$$

$$= -\frac{A}{2} I_T \frac{1}{1 + \cosh Av_{in}(t)} \qquad (22)$$

where

$$A = \frac{q}{kT}$$

$$v_{in}(t) = V_{1o} \cos \omega_{1o} t + V_s \cos \omega_s t$$

V_{1o} = local oscillator voltage

V_s = signal voltage .

For the case we are considering, the amplitude of the signal must be much less than the local oscillator so that transconductance modulation of the μA703 is accomplished essentially by V_{1o}.

Since the displaced cosh function of Equation 22 is an even function, only components which are even multiples of the local oscillator frequency will be present in the output of the μA703. The graphical solution of Equation 22 is shown in Fig. 11 for the first three harmonics. A general algebraic solution is not attempted here, since it would provide little more than exercise. It can be seen from the figure that the large local oscillator drive generates components of the forward transadmittance which are even functions only.

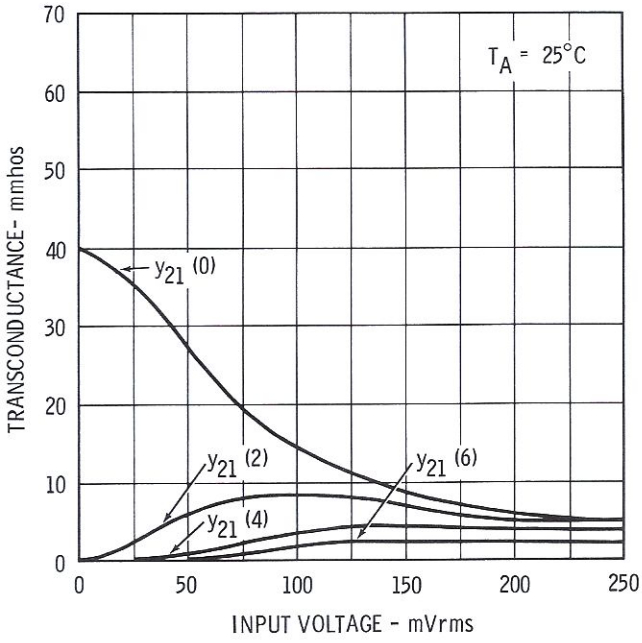

Fig. 11. Forward transconductance of emitter-coupled pair for even-order harmonies.

This "harmonic mixing" is more useful than a conventional mixer because of the lower frequencies and reduced filtering required. For example, it would be much easier to design a 50-MHz oscillator rather than a 100-MHz oscillator to use as a local oscillator. A possible configuration for the mixer is shown in Fig. 12. Reference 5 gives a more detailed description of the μA703 used as a harmonic mixer.

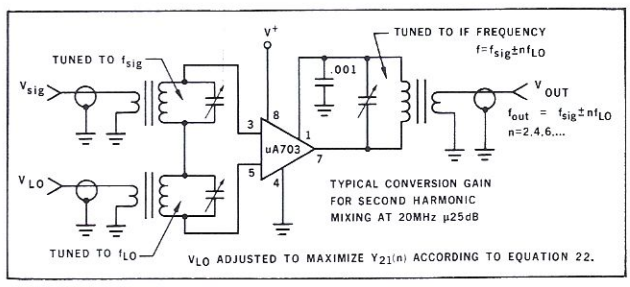

Fig. 12. The μA703 mixer connection.

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APPENDIX

The analysis of the μA703 is simplified if the circuit is initially divided into a differential pair (Q_3 and Q_4 of Fig. 1) and their associated bias circuitry. Relationships may be derived between the transfer function of the pair and its terminal currents and voltages. Following this analysis will be a qualitative discussion of the bias circuitry. Since the signal processing takes place in the emitter-coupled pair, a detailed analysis of the bias network is not necessary. For convenience, Fig. 1 appears again as Fig. A-1; the configuration to be analyzed is shown in Fig. A-2, parts (a) and (b).

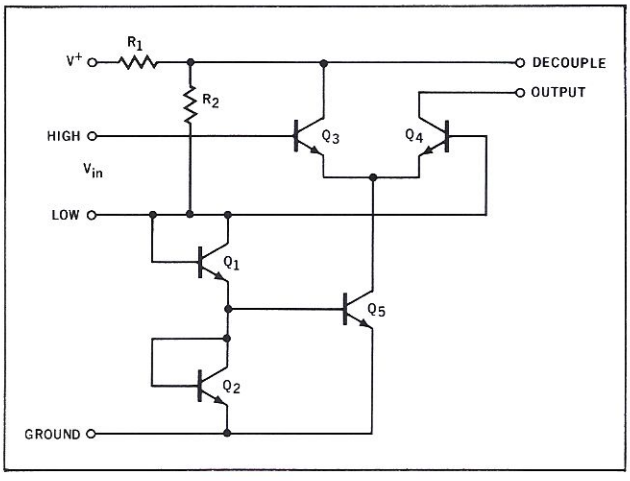


Fig. A-1. Schematic diagram of the μA703.

From elementary device considerations we recall that:

$$I_C = \alpha I_E \cong \alpha I_S \exp qV_{BE}/kT. \qquad (A\text{-}1)$$

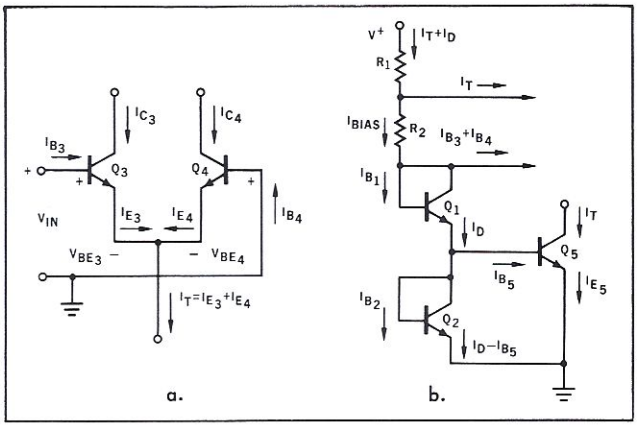


Fig. A-2. μA703 circuit for analysis.

This relates the collector current of a transistor to the base-emitter voltage via an exponential function and a constant I_S, which depends upon the particular geometry and processing used. The alpha term is the familiar common-base current gain of the device.

Manipulation of Equation (A-1) yields

$$V_{BE} = \frac{kT}{q} \ln \frac{I_c}{\alpha I_S}, \qquad (A\text{-}2)$$

and upon substitution of the appropriate quantities, this equation becomes:

$$V_{BE3} = \frac{kT}{q} \ln \frac{I_{C3}}{\alpha_3 I_{S3}}, \qquad (A\text{-}3)$$

$$V_{BE4} = \frac{kT}{q} \ln \frac{I_{C4}}{\alpha_4 I_{S4}}. \qquad (A\text{-}4)$$

From Fig. A-2(a),

$$V_{BE3} - V_{BE4} = Vin \,. \qquad \text{(A-5)}$$

Substitution of Equations A-3 and A-4 in Equation A-5 yields

$$V_{in} = \frac{kT}{q} \ln \frac{I_{C3}}{\alpha_3 I_{S3}} - \ln \frac{I_{C4}}{\alpha_4 I_{S4}}$$

$$= \frac{kT}{q} \ln \frac{I_{C3}}{I_{C4}} + \ln \frac{\alpha_4 I_{S4}}{\alpha_3 I_{S3}} \,. \qquad \text{(A-6)}$$

The second term of Equation A-6 depends upon the degree of matching that exists between the devices Q_3 and Q_4. Assuming reasonably close tolerances, this term is negligible and the input voltage is equivalent to

$$V_{in} = \frac{kT}{q} \ln \frac{I_{C3}}{I_{C4}} \,. \qquad \text{(A-7)}$$

Since the sum of the collector currents of Q_3 and Q_4 is equal to the source current I_T, V_{in} may be expressed in terms of the ratio of total current to output current,

$$V_{in} = \frac{kT}{q} \ln \frac{I_T - I_{C4}}{I_{C4}}$$

$$= \frac{kT}{q} \ln \left[\frac{I_T}{I_{C4}} - 1 \right] . \qquad \text{(A-8)}$$

The forward transadmittance of the μA703 may be obtained from Equation A-8 according to the two-port parameter definition

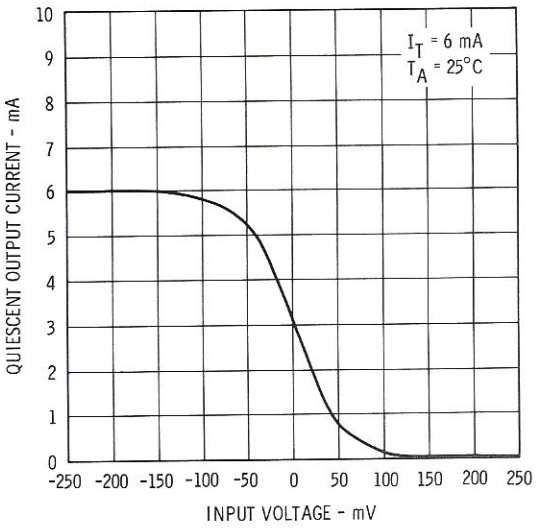
$$y_{21} = \frac{\delta I_o}{\delta V_{in}} \qquad \text{(A-9)}$$

and is

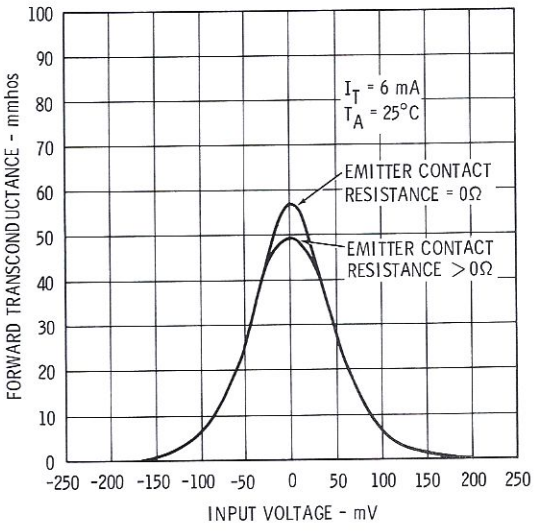
$$y_{21} = 1 / \frac{\delta V_{in}}{\delta I_{C4}}$$

$$= - \frac{qI_{C4}}{kT} \left(1 - \frac{I_{C4}}{I_T} \right) . \qquad \text{(A-10)}$$

Since y_{21} is a function of the ratio of I_{C4} to I_T, Equation A-8 must be solved in order to evaluate y_{21} as a function of V_{in}. The graphical representation of Equation A-8 and A-10 are given in Fig. A-3 (a) and (b). Note that the forward transadmittance, y_{21}, is maximum when the collector currents of the device are equal and that, for matched devices, V_{in} equal to zero corresponds to an output current equal to one-half the source current. A large resistance between the input terminals causes V_{in} to become greater than zero because of the dc offset voltage produced by the finite base currents, I_{B3} and I_{B4}, and results in a shift of the operating point on the y_{21} curve to somewhat below maximum. Therefore, low dc source impedance or transformer coupling should be used with the μA703.



a. Quiescent output current as a function of input voltage.



b. Forward transadmittance as a function of input voltage.

Fig. A-3. Frequency variations in equivalence circuits.

The effect of the emitter contact resistance of Q_3 and Q_4 has not been included in the analysis. By simple superposition it can be shown that contact resistance merely decreases the maximum y_{21} available. The relative position of the maximum with respect to V_{in} is not altered. In addition, some desensitivity to input voltage will be realized because of emitter degeneration.

The total source current, I_T, is approximately equal to the bias current as given by the circuit description following the introduction. Limiting operation of the device is assured if the proper load resistance is used (Eq. 2).

The preceding analysis was based on an ambient temperature of 25 degrees centigrade and a frequency low enough to neglect the effect of the reverse transadmittance of the μA703. If the effect of temperature is included, one will find that output current will vary according to the saturation characteristics of the active devices and the variation of the integrated resistors. Frequency variation of the forward transadmittance will be quite evident above a few megahertz. Curves showing the variation of the two-port parameters of the μA703 are given in Figs. A-4 through A-7, while temperature stability of the quiescent operating point is indicated by Figs. A-8 through A-11. Additionally, quiescent operating conditions as a function of supply voltage are given in Figs. A-12 through A-13.

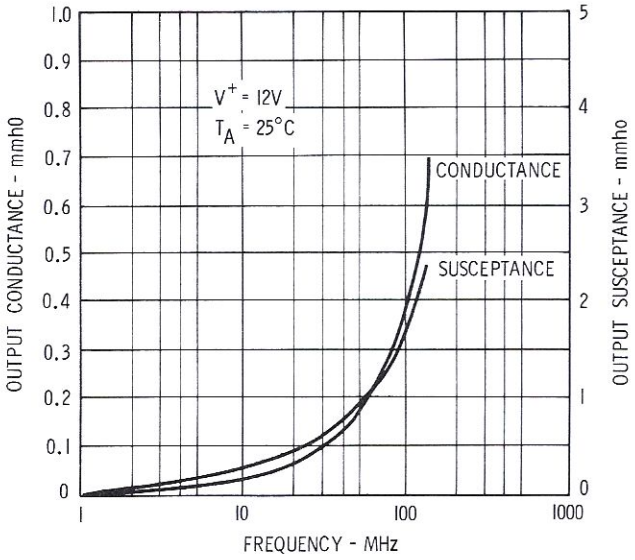


Fig. A-4. Output parameters as a function of frequency.

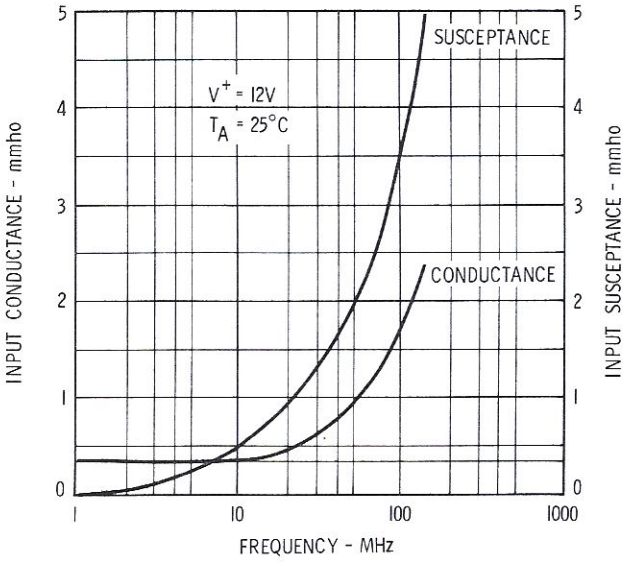


Fig. A-5. Input parameters as a function of frequency.

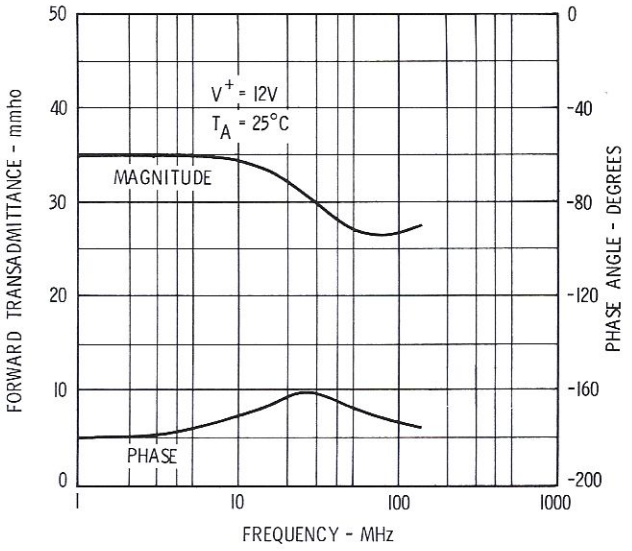


Fig. A-6. Forward transadmittance as a function of frequency.

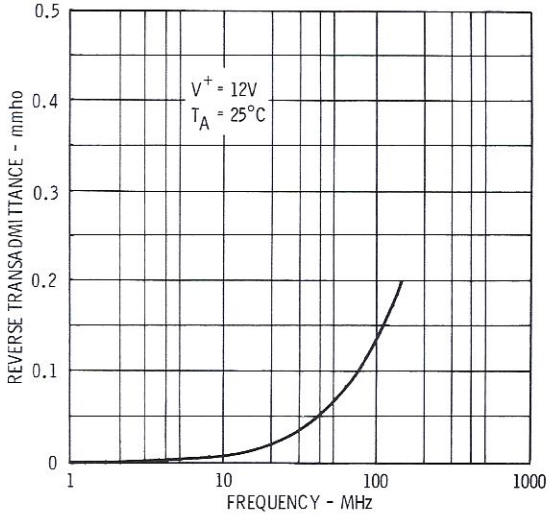


Fig. A-7. Magnitude of reverse transadmittance as a function of frequency.

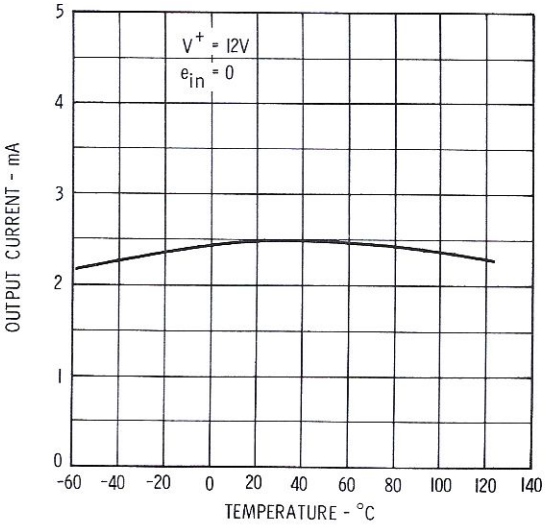


Fig. A-8. Output current as a function of ambient temperature.

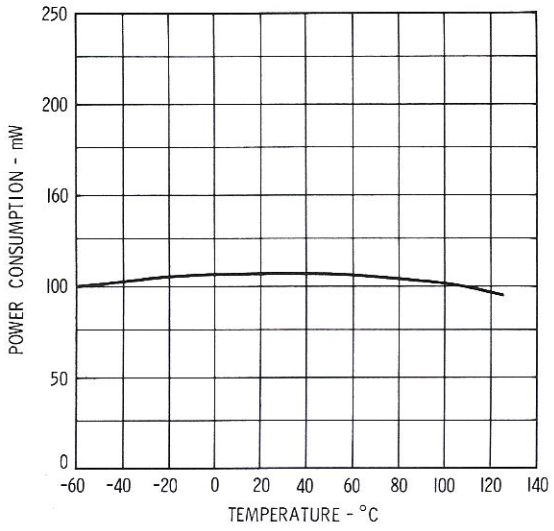


Fig. A-9. Power consumption as a function of ambient temperature.

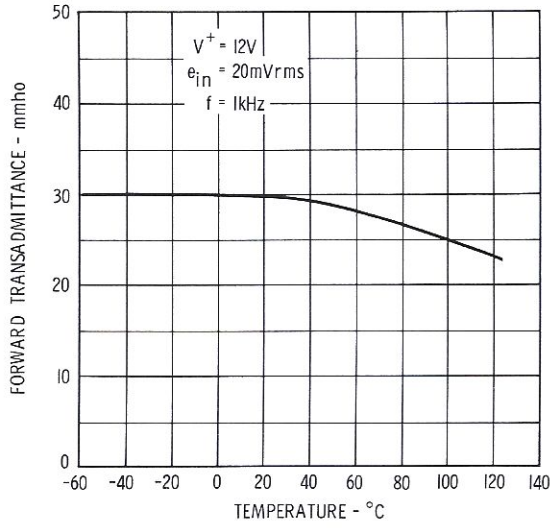


Fig. A-10. Transadmittance as a function of ambient temperature.

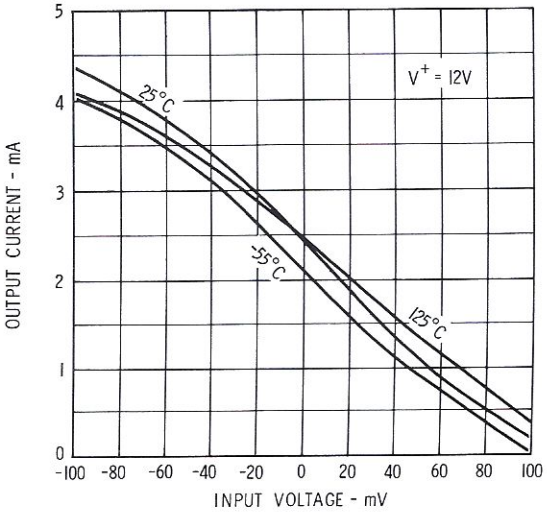


Fig. A-11. Output current as a function of input voltage.

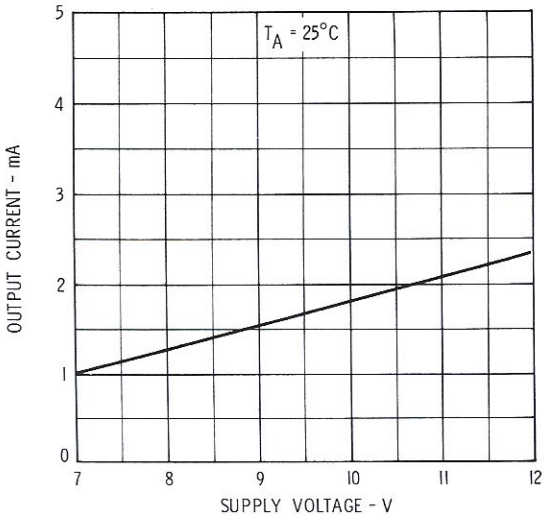


Fig. A-12. Output current as a function of supply voltage.

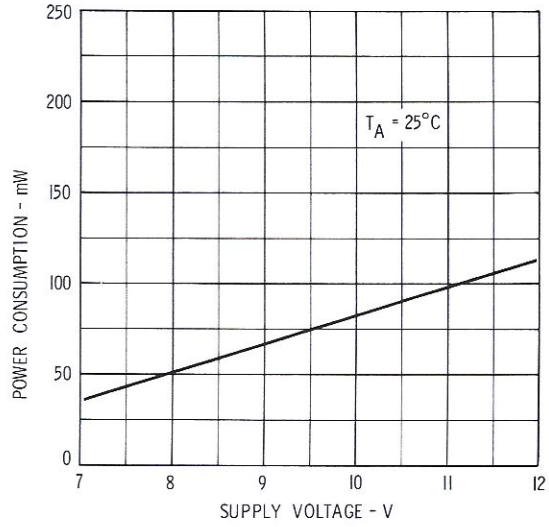


Fig. A-13. Power consumption as a function of supply voltage.

In general, the temperature and supply voltage sensitivities of the μA703 are present because of the voltage variation of biasing diodes, Q_1 and Q_2, and current source, Q_5. A change in the bias current source will cause a proportional change in I_T which will result in a modification of Equations A-1 through A-10. To evaluate, algebraically, the effect of these variations, one need only substitute the appropriate quantities in the above equations.

The frequency variation of the input, output and transfer admittances justify the equivalent circuit of Fig. 3 and are typical of any two-port.